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METHOD FOR DEMODULATING A DIGITAL SIGNAL
SUBJECTED TO MULTIPATH PROPAGATION IMPAIRMENT
AND AN ASSOCIATED RECEIVER

Related Application

This application is based upon prior filed
copending provisional application No. 60/207,028 filed
May 25, 2000, the entire disclosure of which is
5 incorporated herein by reference.

Field of the Invention

The present invention relates to the field of
digital communications, and more particularly, to
demodulation of a serially modulated signal subjected
10 to multipath propagation impairment.

Background of the Invention

A phenomenon in wireless communication
systems, such as digital radio or television
transmission, is multipath propagation. This type of
15 signal degradation occurs when a broadcast signal takes
more than one path from the transmitting antenna to the
receiving antenna so that the receiving antenna
receives multiple signals. One of these multiple
signals may come directly from the transmitting
20 antenna, but several other signals are first reflected
from buildings and other obstructions before reaching

the receiving antenna, and are thus delayed slightly in phase from one another.

The reception of several versions of the same signal shifted in phase results in a composite signal actually being received at the receiving antenna. Two techniques may be used to deal with the multipath propagation of digitally modulated signals. These two techniques are inverse equalization and maximum likelihood sequence estimation (MLSE) detection.

In inverse equalization, an equalizer is implemented, digitally or otherwise, to reverse the propagation effects of multipath on the transmission waveform prior to detection. The equalizer is trained using blind equalization methods, decision feedback methods or by a transmitted training waveform.

There are two fundamental limitations of inverse equalization. The first is the equalizer length, which is a function of the multipath propagation impairment characteristics, namely echo delay and echo amplitude. Equalizer length is necessarily equal to or greater than, and often many times greater than, the multipath delay spread, depending on the amplitude of the multipath pre-echo and/or post-echo components. The second fundamental limitation of inverse equalization is that of 0 dB echo performance. In cases where the amplitudes of delayed signals are equal or nearly equal, the necessary equalizer is usually either unrealizable or impractical.

In MLSE detection systems, a fundamental limitation is complexity. In cases where the channel path count is large and the delay spread is much greater than the symbol interval, the list of survivors becomes unmanageably large, as does the length of the trellis required to represent each survivor. For

example, several MLSE detection systems have been disclosed, such as the ones in Parr et al. (U.S. Patent No. 5,263,026), Polydoros et al. (U.S. Patent No. 5,432,821) and Parr et al. (U.S. Patent No. 5,471,501).

5 In the Parr et al. '026 patent, a method for MLSE demodulation of a received serially modulated signal is disclosed, wherein multipath propagation impairment characteristics are estimated using a least mean square (LMS) algorithm. Rather than converging on
10 an inverse of the multipath propagation impairment, the LMS algorithm converges on an estimate of the multipath propagation impairment. This channel estimate is integrally incorporated into the MLSE algorithm used to determine the symbols making up the serially modulated
15 signal.

 In the Polydoros et al. '821 patent, multipath propagation characteristics are incorporated into the survivor selection process used to accomplish data sequence selection. The survivor selection
20 process is likewise based upon MLSE detection. Also in the Parr et al. '501 patent, MLSE detection is performed using an estimation of the multipath propagation impairment. As discussed above, the MLSE demodulation approach is limited by complexity.

25 A high definition digital television (HDTV) signal is also susceptible to multipath propagation impairment. The HDTV signal is a serially modulated signal based upon the standard set by the Advanced Television System Committee (ATSC) for terrestrial
30 broadcast television in the United States. The ATSC digital television standard was determined by the Grand Alliance and was subsequently accepted by the broadcast community, the consumer electronics industry and the regulatory infrastructure.

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The regulatory infrastructure has mandated a strictly scheduled transition of terrestrial broadcast television in the United States from the National Television System Committee (NTSC) or "analog" standard to the ATSC or "digital" standard. A significant investment is in place on behalf of the broadcast industry to support this planned transition. Similarly, many consumers have purchased ATSC television receiver equipment that include new ATSC system compliant DTV television sets and DTV television set-top converters.

However, the ATSC standard, in its present form, is deficient in its susceptibility to multipath propagation impairment. In side-by-side comparisons, ATSC reception, i.e., the new digital system, is often inferior to NTSC reception, i.e., the conventional analog system. Additionally, ATSC mobile reception suffers substantially more degradation due to multipath propagation impairment than NTSC mobile reception. Signal strength and signal-to-noise (SNR) ratios are typically not at issue, as unanticipated inferior reception manifests itself at high levels of received signal power and at high receiver SNR ratios. This fact, coupled with spectral analysis of received ATSC DTV signals, points directly to multipath propagation impairment as the cause of the inferior reception.

Various efforts have been made in the area of DTV reception. For example, Park et al. (U.S. Patent No. 5,592,235) discloses combining reception, appropriate to terrestrial broadcast and to cable broadcast, both in a single receiver. Also included in these various efforts is Oshima (U.S. Patent No. 5,802,241), which discloses a plurality of modulation components modulated by a plurality of signal components. Both of these references disclose the use

of equalization. As discussed above, complexity of an equalizer is a fundamental limitation.

With respect to enabling the initial acquisition of digitally modulated signals that are
5 severely distorted by multipath propagation impairment, decision-feedback equalizers (DFE) are not suitable. For this purpose, a reference or training waveform is typically introduced. The use of a reference sequence equalizer for equalizing GA-HDTV signals is disclosed
10 in Lee (U.S. Patent No. 5,886,748). Unfortunately, the Lee '748 patent does not overcome the limitations associated with inverse channel equalizers.

Summary of the Invention

In view of the foregoing background, it is
15 therefore an object of the present invention to provide a method for demodulating a received digitally modulated signal that is subjected to multipath propagation impairment, particularly when multiple signals of the received signal defining the multipath
20 propagation impairment are substantially equal to one another.

Another object of the present invention is to provide a corresponding digital receiver that is relatively straightforward to implement for
25 demodulating the received digitally modulated signal.

These and other objects, advantages and features in accordance with the present invention are provided by a method for demodulating a received digitally modulated signal subjected to multipath
30 propagation impairment. The method preferably comprises estimating the multipath propagation impairment of the received digitally modulated signal using a channel estimator, and estimating at least one

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symbol of the received digitally modulated signal using a symbol estimator.

The method preferably further includes adjusting the at least one estimated symbol based upon
5 the estimated multipath propagation impairment to generate an estimate of the at least one symbol as impaired by the multipath propagation, and at least one error signal is generated by comparing the estimate of the at least one symbol as impaired by the multipath
10 propagation to the received digitally modulated signal. The at least one error signal is then preferably used for estimating remaining symbols to be demodulated.

The method preferably further comprises using the at least one error signal for refining the
15 estimated multipath propagation impairment. Next, the method also preferably further comprises estimating at least one next symbol, and adjusting the estimate of the at least one next symbol based upon the refined estimated multipath propagation impairment for
20 generating an estimate of the at least one next symbol as impaired by the multipath propagation.

The at least one error signal is preferably refined by comparing the estimate of the at least one next symbol as impaired by the multipath propagation to
25 the received digitally modulated signal. Refining the at least one error signal preferably further comprises comparing the estimate of the at least one next symbol as impaired by the multipath propagation to the at least one error signal resulting from at least one
30 previous comparison.

Estimating the multipath propagation impairment may be based upon an adaptive algorithm, or based upon a training waveform embedded in the received digitally modulated signal. Similarly, estimating the
35 at least one symbol may be based upon an adaptive

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algorithm, or based upon the training waveform embedded in the received digitally modulated signal. With respect to the adaptive algorithms, each algorithm may comprise a respective least mean square (LMS) algorithm
5 that has applied thereto a convergence coefficient. The convergence coefficient is preferably based upon the received digitally modulated signal.

After the at least one symbol has been estimated, the remaining symbols to be demodulated are
10 preferably estimated based upon linear estimation. This is performed based upon the at least one error signal. In other words, linear estimation of the remaining symbols or adaptive estimation of the remaining symbols allows the received digitally
15 modulated signal to be demodulated when impaired by multichannel propagation, particularly when multiple signals of the received signal defining the multipath propagation impairment are substantially equal to one another.

20 Since possible combinations of the symbols to be demodulated are preferably not estimated, as is typically the case for a MLSE equalizer, the complexity of a digital receiver demodulating the received digital signal is minimized. Consequently, performing an
25 adaptive estimation or a linear estimation for the symbols to be demodulated overcomes the limitations applicable to inverse equalization and MLSE estimation, as discussed in the background section.

The received digitally modulated signal
30 preferably comprises at least one of a digital broadcast television signal, a digital broadcast radio signal, a digital cellular telephone signal, and a digital wireless local area network (LAN) signal. Of course, the method according to the present invention
35 may also be applied to other radio systems and to

communication through various types of media. In addition, the received digitally modulated signal may be a digitally serial modulated signal.

Another aspect of the invention is directed
5 to a method for simultaneously demodulating a plurality of received digitally modulated signals subjected to multipath propagation impairments. The method preferably comprises estimating the multipath propagation impairments of the plurality of received
10 digitally modulated signals using a plurality of channel estimators, and estimating at least one symbol of each of the plurality of received digitally modulated signals using a plurality of symbol estimators.

15 Each estimated symbol is preferably adjusted based upon the corresponding estimated multipath propagation impairment to generate an estimate of each symbol as impaired by the corresponding multipath propagation, and at least one error signal is
20 preferably generated by comparing a summation of the estimates of the symbols as impaired by the corresponding multipath propagation to the plurality of received digitally modulated signals. The at least one error signal is preferably used for estimating
25 remaining symbols of each of the plurality of received digitally modulated signals to be demodulated.

Another aspect of the present invention is directed to a receiver for demodulating a received
30 digitally modulated signal subjected to multipath propagation impairment. The digital receiver preferably comprises a channel estimator for estimating the multipath propagation impairment of the received digitally modulated signal, and a symbol estimator
35 connected to the channel estimator for estimating at

least one symbol of the received digitally modulated signal.

The channel estimator preferably adjusts the at least one estimated symbol based upon the estimated
5 multipath propagation impairment to generate an estimate of the at least one symbol as impaired by the multipath propagation. The digital receiver may further comprise a summing network connected to the channel estimator and to the symbol estimator for
10 generating at least one error signal by comparing the estimate of the at least one symbol as impaired by the multipath propagation to the received digitally modulated signal. The symbol estimator preferably uses the at least one error signal for estimating the
15 remaining symbols to be demodulated.

Brief Description of the Drawings

FIG. 1 is a simplified block diagram of a digital transmitter including a continuous-time modulator and a channel model in accordance with the
20 prior art.

FIG. 2 is an illustration of a segment of a digitally modulated waveform comprising a plurality of symbols in accordance with the prior art.

FIG. 3 is an illustration of various physical
25 objects providing propagation paths for a transmitted signal in accordance with the prior art.

FIG. 4 is an illustration of a five-signal multipath model being applied to a digitally modulated signal in accordance with the prior art.

30 FIG. 5 is a simplified block diagram of a digital transmitter including a time-sampled modulator and a channel model in accordance with the prior art.

FIG. 6 is a block diagram on the architecture of a digital receiver based upon equalization in accordance with the prior art.

FIG. 7 is an illustration of a two-signal
5 multipath model having a benign multipath being applied to a digitally modulated signal in accordance with the prior art.

FIG. 8 is an illustration of the successful equalization of a received signal impaired by a
10 moderate two-signal multipath model in accordance with the prior art.

FIG. 9 is an illustration of the failure of conventional equalization when a received signal impaired by a severe two-signal multipath model is
15 applied thereto in accordance with the prior art.

FIG. 10 is an illustration on the 0 dB echo problem, both static and dynamic, to conventional equalizers in accordance with the prior art.

FIG. 11 is a flow diagram for demodulating a
20 received digitally modulated signal in accordance with the present invention.

FIG. 12 is a simplified block diagram of a digital receiver illustrating the cooperation between symbol estimation and channel estimation in accordance
25 with the present invention.

FIGS. 13-16 are illustrations of demodulation of the first six symbols of a received signal impaired by multipath propagation, with the demodulation based upon linear estimation in accordance
30 with the present invention.

FIG. 17 is a block diagram of a digital receiver having adaptive channel estimation in accordance with the present invention.

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FIG. 18 is a block diagram of a digital receiver having adaptive symbol estimation in accordance with the present invention.

FIG. 19 is a block diagram illustrating a
5 digital receiver having joint adaptive channel estimation and symbol estimation in accordance with the present invention.

FIG. 20 is a detailed block diagram of a
10 digital receiver having joint adaptive channel estimation and symbol estimation associated with a plurality of independent modulation sources in accordance with the present invention.

Detailed Description of the Preferred Embodiments

The present invention will now be described
15 more fully hereinafter with reference to the accompanying drawings, in which preferred embodiments of the invention are shown. This invention may, however, be embodied in many different forms and should not be construed as limited to the embodiments set
20 forth herein. Rather, these embodiments are provided so that this disclosure will be thorough and complete, and will fully convey the scope of the invention to those skilled in the art. Like numbers refer to like elements throughout and prime and multiple prime
25 notations are used in alternate embodiments. The dimensions of layers and regions may be exaggerated in the figures for greater clarity.

Referring initially to FIGS. 1-10, a digital transmitter and a digital receiver of the prior will be
30 discussed, including the impact of multipath propagation on a digitally modulated signal. A simplified block diagram of a digital transmitter 10 including a continuous-time modulator 12 and a channel modeler 14 is illustrated in FIG. 1.

In the digital transmitter and channel model
10, $x(n)$ represents the data sequence applied to a
modulator 12, which generates a modulated waveform $s(t)$
represented in real time. The modulated waveform is
5 broadcast through a propagation channel 14 having a
time response $h(t, \tau)$ in a convolutional continuous-time
domain τ that varies continuously over time t . Noise
 $n(t)$ is added via a summing network 16 for generating
the resulting waveform, which is represented by the
10 waveform $r(t)$.

Referring now to FIG. 2, a modulated waveform
18, for example, comprises a series of different
modulation symbols or symbols 18₁-18₆. The modulation
symbols 18₁-18₆ may also be referred to simply as
15 symbols. Each symbol is selected from an ensemble of
unique shapes, i.e., of varying amplitudes and phases.
Each unique shape represents a digital state or group
of digital information bits.

These symbols 18₁-18₆ are transmitted
20 serially, i.e., one right after the other. Digital
serial modulation is contrasted with Orthogonal
Frequency Division Multiplexing (OFDM/COFDM) in that
serial modulation carries information serially while
OFDM/COFDM carries information both serially and across
25 the modulation spectrum. Although OFDM/COFDM can offer
multipath propagation advantages, digital serial
modulation is superior in that it is simpler and does
not suffer from distortion due to extreme ratios of
peak-to-average power.

30 In an ideal world, digital transmission
passes through a medium, such as air or space, in a

straight line and unimpaired. As illustrated in FIG. 3, a transmitter 30 transmits a signal via transmit antenna 32 to receive antenna 34, which is connected to a receiver 36. Realistically, however, the transmitted signal is subjected to obstacles. The transmitted signal is reflected from objects such as buildings 20, bridges 22, aircraft 24, and other man-made and natural structures or obstacles 26. Consequently, the transmitted signal arrives at the receiver 36 after having passed through any number of multiple paths, such as any one of the five paths illustrated in FIG. 3.

For a clear path, a clear channel response may be represented as a single signal component 40_1 , as illustrated in FIG. 4. The single signal component 40_1 indicates a single time of arrival (TOA), with time progressing from left to right. A multipath response is indicated by multiple signal components 40_1 - 40_5 , with each signal component indicating a different arrival time, a different amplitude and a different phase which may be either positive or negative. In the illustrated example, there are five paths in the transmission medium at some instant in time. Each signal component corresponds to one of these paths.

Assuming that a single symbol 18_1 from the received signal travels across a single path, then it is received at a single arrival time as part of signal component 40_1 . This arrival time corresponds to a single delay and at the amplitude and phase associated with the first single path. However, in a multipath situation, the second path contributes a component 18_{1a} to the received signal (i.e., there is another symbol 18_1 provided by multipath signal component 40_2) at a second delay with a second associated amplitude and

phase. Likewise the third path contributes a component 18_{1b} to the received signal (i.e., there is another symbol 18_1 provided by multipath signal component 40_3), this time with a negative phase. Similarly, the fourth and fifth paths each contribute a component 18_{1c} , 18_{1d} to the received signal (i.e., there are two more symbols 18_1 provided by multipath signal components 40_4 and 40_5).

What the receiver 36 sees is the sum of these five multipath components, as represented by signal 50, which is distorted compared to the original transmitted symbol 18_1 . The assumption is now made that an entire digital serial modulation waveform is transmitted 18 to include six consecutive symbols 18_1 - 18_6 . In this case, the received signal 19 is distorted by the presence of five separate paths in such a way as to cause the signal 18 to interfere with itself. The received signal 19 is unrecognizable in this case due to the impairment by the multipath propagation.

When the receiver and demodulation techniques are implemented digitally, digital equalization and multipath analysis lend themselves to sampled-time digital modeling and analysis. As such, the digital transmitter and channel estimate 10' illustrated in FIG. 5 includes a modulator 12' and a channel modeler 14', which are represented in sampled-time as compared to continuous-time shown in FIG. 1. In this illustration, continuous-time t is replaced by time-sampling index n and the continuous convolutional-time domain τ is replaced by the time-sampling convolutional index m .

This model allows for complex (real and imaginary) signal representation and for time sampling

intervals which may be integer fractions of the symbol interval. In this model, the same transmission data sequence $x(n)$ as that in FIG. 1 is applied to a time-sampling digital modulator 12' yielding the time-sampled modulated waveform $s(n)$. The time-sampled digitally modulated waveform $s(n)$ is applied to the time-sampled channel model $\bar{h}(n,m)$ 14', which is made up of a sequence of time-sampled impulse responses in index m , one per time index n .

Time-sampled noise $n(n)$ is added via a summing network 16 to the output of the time-sampled channel or multipath model process $\bar{h}(n,m)$ to yield a time-sampled representation of the received modulated waveform $r(n)$, again in time index n . Successful demodulation requires sufficient consideration of channel distortion $\bar{h}(n,m)$ in the process of estimating the modulation data sequence $x(n)$.

Referring now to FIG. 6, an equalization process or method for a digital receiver 60' will be discussed. An equalizer 62 is connected to a demodulator 64. An approximation $\hat{h}^{-1}(n,m)$ to the inverse $\bar{h}^{-1}(n,m)$ of the channel response $\bar{h}(n,m)$ is applied to the received waveform $r(n)$. The resulting output $\hat{s}(n)$ is an estimate of the original modulation waveform $s(n)$. The demodulator 64 operates on the modulation waveform

estimate $\hat{s}(n)$ to produce an estimate $\hat{x}(n)$ of the modulation data sequence $x(n)$.

Provided that the channel-inverse equalization response $\overline{h^{-1}}(n,m)$ exists and can be

5 approximated sufficiently as $\hat{\overline{h^{-1}}}(n,m)$ within practical implementation limitations, such as finite impulse response (FIR) filter duration and resolution, the output $\hat{x}(n)$ of the demodulator 64 will be a sufficiently accurate reproduction of the modulation data sequence

10 $x(n)$. However, equalizer length, equalizer tap resolution and the existence and/or practical implementation of the inverse channel response are factors that effect the practical implementation of the equalization process.

15 The operation and consequent limitations of conventional equalizer techniques will now be described with an example. A straightforward example of digital equalization based on a two-signal multipath channel is illustrated with reference to FIG. 7. Again,

20 one starts with a clear path which exhibits the clear channel response. The single signal component 40₁ indicates the first single TOA, again with time progressing from left to right.

In the two-signal multipath response, each

25 signal indicates a different arrival time with a different amplitude and phase. Here we show two signal components 40₁ and 40₂, with each signal component corresponding to one of two propagation paths in this example. Assuming a single symbol 18₁ travels across a

single path, it is received at a single arrival time corresponding to a single delay and at the amplitude and phase associated with the first single path 40_1 .

In a two-signal multipath situation, the
5 second path contributes a second component 18_{1A} to the received signal (i.e., there is another symbol 18_1 provided by multipath signal component 40_2) at the second delay with a second associated amplitude and phase. What the receiver 36 sees is the sum (signal
10 $50'$) of these two multipath components which is distorted compared to the original transmitted symbol.

An assumption is now made that an entire digital serial modulation waveform 18 is transmitted to include the six consecutive symbols 18_1-18_6 . In this
15 case, the received signal $19'$ is distorted by the presence of two distinct paths in such a way as to cause the signal to interfere with itself. The received signal $19'$ is severely distorted when compared to the original modulated signal 18.

20 Currently, receivers compensate for this multipath propagation impairment, i.e., distortion, using equalization techniques. Considering the same transmitted serial modulated waveform 18, along with the same two-signal multipath response example as
25 discussed above with reference to FIG. 7, the received signal $19'$ as shown earlier is again shown with reference to FIG. 8.

Equalization, as readily understood by one skilled in the art, employs a finite impulse filter
30 (FIR) 62 for the received signal $19'$, which is assumed to have a dominant primary path component 40_1 . This filter (or equalizer) 62 operates on the principle of adding delayed versions of the received signal so as to cancel non-primary paths of lesser strength.

In the example illustrated in FIG. 8, the equalizer begins by introducing a delayed component 70_1 to the primary received signal component 70_0 . The delayed signal component 70_1 is designed to cancel the secondary multipath component 40_2 , which is smaller in amplitude with respect to the primary multipath component 40_1 . The result is a signal 21 with most of the multipath distortion cancelled.

However, there is still some residual distortion at twice the echo delay. So the equalizer is adjusted by adding a tap 70_2 , this time to cancel the compound echo at twice the path delay. The result is a much cleaner signal, as illustrated by signal $21'$. This may be repeated with two more taps $70_{3,4}$ to produce an even cleaner signal $21''$. The resulting equalized waveform $21''$ is very clean, almost indistinguishable from the modulated waveform 18 .

Unfortunately, equalization may not be sufficient when the echo is almost as strong as the direct path signal. Referring now to FIGS. 9 and 10, the two-signal 40_1 and 40_2 multipath scenario will be addressed again, except this time multipath signal component 40_2 is almost as strong as the direct signal component 40_1 . Each signal component 40_1 and 40_2 corresponds to one of two propagation paths.

With a six-tap 70_{0-5} equalizer, the resulting signal 25 has the multipath propagation impairment cancelled at the echo and out to four compound echoes. However, there is a great deal of residual noise, not evident on the left, where cancellation is illustrated, but on the right, where the compound echos go uncanceled. This example is carried out to a nine-tap 70_{0-8} equalizer which passes the received signal 23 (first tap 70_0), cancels the channel echo (second tap

70₁) and cancels seven subsequent compound echoes 70₂₋₈, out to 8 times the original path delay, as indicated by signal 25'.

The result again shows cancellation on the left, but there is still significant noise remaining, as indicated on the right. However, a more realistic picture of what is happening is made available when one adds the effect of the multipath and the equalizer on the symbols arriving before the six 18_{1-18₆}, that are illustrated in the digital serial modulated waveform 18. The resulting signal 27 is as bad as, if not worse, than the original received waveform 23.

The equalization process has another problem with respect to the 0-dB echo, as illustrated with reference to FIG. 10. Considering the multipath profile where two signal components 40₁ and 40₂ are very close in amplitude, with the first signal component 40₁ dominating. The necessary equalizer response would be a "post" equalizer, which cancels the second component 40₂ with respect to the first component 40₁.

Suppose now that the multipath response were to change, and the second signal component 40_{2B} began instead to dominate the first signal component 40_{1B}. This is because the first signal component suffered attenuation, or because the first signal was blocked and both paths represent reflections. In this case, the multipath cancellation requires a "pre" equalizer filter, cancelling the first signal component 40₁ to arrive with respect to the second signal component 40₂.

As discussed in the background section, these equalizers are long, much longer than their corresponding path delays. This characteristic makes them difficult to implement. As a practical matter, each additional required equalizer tap introduces

additional noise into the system. The more taps, the more difficult it is to demodulate, even when the equalizer can implement all the taps. The discontinuity from the "post" equalizer to the "pre" equalizer represents a very difficult equalizer training problem. When the multipath response has two equal signal components 40_{1A} and 40_{2A} , equalization can not be used.

The present invention will now be described with reference to FIGS. 11-20. Referring to the flow chart illustrated in FIG. 11, from the start (Block 90) the method for demodulating a received digitally modulated signal that is subjected to multipath propagation impairment comprises estimating the multipath propagation impairment of the received digitally modulated signal using a channel estimator at Block 92, and estimating at least one symbol of the received digitally modulated signal using a symbol estimator Block 94.

The method further includes adjusting the at least one estimated symbol based upon the estimated multipath propagation impairment to generate an estimate of the at least one symbol as impaired by the multipath propagation Block 96, and at least one error signal is generated by comparing the estimate of the at least one symbol as impaired by the multipath propagation to the received digitally modulated signal at Block 98. In other words, the initial symbol sequence estimate is convolved with the multipath estimate, and the result of the convolution is subtracted from the received signal to generate the at least one error signal. The at least one error signal is then preferably used for estimating remaining

symbols to be demodulated at Block 100, and the method may be stopped at Block 102.

5 The method according to this embodiment of the present invention advantageously combines channel estimation and symbol estimation for demodulating the received digitally modulated signal, which may be serial. This avoids the limitations inherently associated with inverse equalization and MLSE detection as discussed above. The method may be used to
10 successfully demodulate in the presence of all the multipath profiles that can be corrected with an equalizer. In addition, the received signal may also be successfully demodulated in the presence of all the multipath profiles that can not be corrected with an
15 equalizer without requiring extremely long processing for multiple compound delays, or without requiring special processing to accommodate discontinuities as required by the equalizer. In other words, "killer" equalizer tracking problems are avoided with the method
20 according to the present invention. There is also an increased signal-to-noise ratio advantage in the present invention due to a reduction of required taps.

The present invention thus overcomes the dilemma of implementing a possibly non-existent
25 inverse-channel response and reduces the resolution required of the associated processing with respect to that required of comparable channel-inverse equalization techniques.

Referring now to the digital receiver 120
30 illustrated in FIG. 12, the two parts include symbol estimation using a symbol estimator 122 and multipath estimation using a channel estimator 124. Initial multipath estimation may be as straightforward as correlating against a reference sequence like an a-

priori PN sequence, as readily understood by one skilled in the art, whereas symbol estimation can be as straight-forward as linear combination or demodulation of the error vector, as also readily understood by one skilled in the art.

Cooperative channel estimating demodulation will first be discussed. The serial modulated waveform 18 used in previous examples will again be the center point of the discussion. In addition, the five-path multipath profile 40₁₋₅ shown earlier will also be the center point of the discussion.

The received signal 19 is stored in a memory 126. Suppose one could determine or at least estimate what the multipath profile looked like 130₁₋₅ by estimating the relative delay, amplitude and phase of every path. Suppose also that one could search for or recognize the first symbol 18₁ in the received waveform 19.

Then, knowing the multipath profile 130₁₋₅ or at least having a good appreciation as indicated by signal 132, one could assess the effects of this multipath profile on the first symbol 131, as illustrated in FIG. 13. By subtracting this multipath-corrupted first symbol 18₁ from the received waveform 19 using a summing network 128, one gets an error signal 134.

In actuality, the first symbol 18₁ was recognized above by choosing the symbol 131 which minimized this error waveform 134. We continue to demodulate this same serial modulated waveform 18. We already know the first symbol 18₁, and we are working off of the error signal 134 derived from the previous

step, and we have a good estimate of the multipath response **130₁₋₅**.

In fact, we use the first symbol **18₁** to refine our good estimate of the multipath response and make it better. The next step is to estimate the second symbol **137** again by driving the estimation process, which causes convergence of the error signal **134** to a set level, such as zero.

Application (e.g., convolution) to the multipath estimate yields an estimate **138** of the component of the received waveform which corresponds to the second symbol **18₂**. Subtraction yields a new error signal **140**, which is closer to flatline than the previous error signal **134**, as illustrated in FIG. 14. This means we are making progress and that we are heading in the right direction.

Referring to FIG. 15, the same transmitted serial modulation waveform **18** is offered as a reference. We already know the first two symbols **18₁** and **18₂**, and we are working off of the new error signal **140** from the previous step. We have a good estimate of the multipath response **130₁₋₅**, which is again refined with the benefit of the error signal **140** based upon the previously demodulated symbol.

The next step is to estimate the third symbol **141**, again by driving the error signal **140** to zero. The resulting error signal **142** is shown next, which incorporates the effects of multipath, as estimated, on the demodulated third symbol **18₃**. After using the third symbol **18₃** and the new error signal **142** to again update the multipath estimate **130₁₋₅**, the fourth symbol **146** is estimated. A new error signal **148** is generated.

Again, the same transmitted serial modulation waveform 18 is offered as a reference. We already know the first four symbols 18₁-18₄ from earlier in the process. We are working off of the new error signal 148 from the previous step. Again, we have a good estimate of the multipath response 130₁₋₅, again refined using the new error signal 148 and the fourth symbol 18₄, just demodulated.

The next step is to estimate the fifth symbol 149, again by driving the error signal 148 to zero. The resulting error signal 150 is shown next, which again incorporates the effects of multipath, as estimated, on this newest demodulated symbol. After using the fifth symbol 18₅ and the new error signal 150 to again update the multipath estimate, the last symbol 152 is estimated, and a new error signal 154 is generated.

The flatline of error signal 154 indicates successful demodulation, as illustrated in FIG. 16. Any deviation at this point from zero would be due to one or more of the following causes. Noise in the received signal; errors in the multipath estimate, which is normal in noisy channels but limited with respect to equalizer tap noise due to the absence of compound equalizer echos; and demodulation errors, which are expected when operating near the SNR threshold which is much lower than that experienced by equalizer-based systems in severe multipath environments. Any error left can be used to drive an adaptive multipath or channel estimation.

In another embodiment of the digital receiver, adaptive algorithms are applied to both processes, i.e., channel estimation and symbol estimation. The first part of this method is an

adaptive channel estimation process illustrated in FIG. 17. In this digital receiver 120', the received signal waveform $r(n)$ is stored in the memory 126 as a received signal vector $\bar{r}(n,k)$ whose depth is represented by index

5 k . An adaptive algorithm 170 may be part of the channel estimator 172. It is assumed that the transmission modulation waveform $s(n)$ is known and stored as a vector $\bar{s}(n,k)$ also indexed in depth by sample index k . The following convention applies to

10 each element of the transmission modulation-waveform vector $\bar{s}(n,k)$:

$$s(n,k) = s(n+k)$$

This same convention applies to all vector variables using (n,k) arguments throughout this

15 document. The vector modulation waveform $\bar{s}(n,k)$ is applied to an estimate $\hat{h}(n,m)$ of the transmission-channel sampled-time response $\bar{h}(n,m)$. For purposes of initialization, the transmission-channel sampled-time response-estimate $\hat{h}(n,m)$ may be initialized, at the

20 beginning of the process, to unity-gain at $m=0$ and zero response at all other values of m .

When the vector modulation waveform $\bar{s}(n,k)$ is applied to the channel-response estimate $\hat{h}(n,m)$, the

result is an estimate vector $\hat{\bar{r}}(n,k)$ of the corresponding received waveform vector $\bar{r}(n,k)$. These two vectors are subtracted in the summing network 128, resulting in the error signal vector $\bar{e}(n,k) = \hat{\bar{r}}(n,k) - \bar{r}(n,k)$. This error

5 signal drives the adaptation process 170, which modifies the channel-response estimate $\hat{h}(n,m)$ in the channel estimator 172 in such a manner as to cause the error vector $\bar{e}(n,k)$ to converge on the corresponding zero vector.

10 Any number of adaptive algorithms may be used to gradually modify the channel response vector estimate $\hat{h}(n,m)$ towards a successively more accurate representation of the channel response vector $\bar{h}(n,m)$. The LMS algorithm is known for its advantages in

15 tracking non-stationary processes and is used, for that reason, as an example. The LMS algorithm requires a convergence coefficient μ . In this case, the convergence coefficient is defined at every time-sample point n over the vector depth index k . The vector

20 convergence coefficient is denoted $\bar{\mu}_h(n,k)$. An LMS adaptation recursion equation suitable for adaptation at every time sample n is

$$\hat{h}(n+1,m) = \hat{h}(n,m) - \sum_{k=k_{\min}+m_{\max}}^{k_{\max}+m_{\min}} \mu_h(n,k-m) e(n,k) s(n,k-m)$$

An advantageous feature of the present invention is contained in the second part of this method, which is the progressive adaptive estimation of the transmission modulation waveform $s(n)$. An adaptive
5 algorithm 180 may be part of the symbol estimator 172. As best illustrated by the digital receiver 120" in FIG. 17, an adaptive process 180 is used to converge on the most likely modulation waveform when the channel response approximation vector $\hat{h}(n,m)$ in the channel
10 estimator 172' is sufficiently known to be a sufficiently valid approximation of the channel response vector $\bar{h}(n,m)$.

In this digital receiver 120", the received signal waveform $r(n)$ is again stored in the memory 126
15 as a received signal vector $\bar{r}(n,k)$, whose depth is represented by index k . It is assumed that the channel response vector $\bar{h}(n,m)$ is sufficiently known and stored as a vector $\hat{h}(n,m)$ also indexed in depth by sample index k . An estimate $\hat{s}(n,k)$ of the vector modulation waveform
20 $\bar{s}(n,k)$ is applied to the stored channel time-response vector-estimate $\hat{h}(n,m)$. For purposes of initialization, the estimate $\hat{s}(n,k)$ of the transmitted modulation waveform may be initialized, at the beginning of the process, to all zeroes.

When the vector modulation-waveform approximation $\hat{s}(n,k)$ is applied to the channel-response estimate $\hat{h}(n,m)$, the result is an estimate vector $\hat{r}(n,k)$ of the corresponding received waveform vector $\bar{r}(n,k)$.

5 These two vectors are subtracted in the summing network 128, resulting in the an error signal vector

$\bar{e}(n,k) = \hat{r}(n,k) - \bar{r}(n,k)$. This error signal drives the

adaptation process, which modifies the estimate $\hat{s}(n,k)$ of the vector modulation waveform $\bar{s}(n,k)$ in such a
10 manner as to cause the error vector $\bar{e}(n,k)$ to converge on the corresponding zero vector.

Again, any number of adaptive algorithms may be used to gradually modify vector modulation waveform approximation vector $\hat{s}(n,k)$ towards a successively more
15 accurate reproduction of the transmitted modulation waveform vector $\bar{s}(n,k)$. Again, the LMS algorithm is known for its advantages in tracking non-stationary processes and is used, for that reason, as an example. The LMS algorithm requires a convergence coefficient
20 μ . In this case, the convergence coefficient is defined at every time-sample point n over the vector depth index k . The vector convergence coefficient is denoted $\bar{\mu}_i(n,k)$. An LMS adaptation recursion equation suitable for adaptation at every time sample n is

$$\hat{s}(n+1, k-1) = \hat{s}(n, k) - \sum_{m=m_{\min}}^{m_{\max}} \mu_s(n, k) e(n, k-m) \hat{h}(n, m)$$

The process is completed through the selection of a suitable delay index k_d from which to generate a modulation waveform estimate $\hat{s}(n+k_d)$ suitable
5 for demodulation through demodulator 184. This demodulation process yields an estimate $\hat{x}(n+k_d)$ of the original corresponding data sequence element $x(n+k_d)$.

What has just been described is a method of adaptively converging on an estimate $\hat{s}(n)$ of the
10 modulation waveform $s(n)$. However, many serial data-modulation processes are linear. In each of these cases, an appropriate substitution of variables serves to convert this method into an equivalent form where adaptation is applied directly to an estimate $\hat{x}(n)$ of
15 the modulation data-sequence $x(n)$.

An example of such a system where this is possible is the 8-VSB modulation applicable to the ATSC standard for terrestrial television broadcast. Such direct estimation of the modulation data-sequence
20 results in a significant advantage in computational efficiency. Such direct estimation of the modulation data-sequence through the substitution described is also relevant and applicable to the remainder of this disclosure.

Further savings in computational efficiency may be realized by considering the restrictions on modulation symbol-states associated with a modulation data-sequence $x(n)$ specific to a given modulation system

5 in question. Again, referring to the 8-VSB ATSC DTV example, the modulation data-sequence in this case is limited to 8 states (four positive states and four negative states, namely: -7, -5, -3, -1, 1, 3, 5 and 7).

10 An improvement in bit-error-rate (BER) performance is achievable as follows. In many modulation systems, linear modulation applies and forward error correction is employed, whether by trellis coded modulation, other convolutional coding or
15 by block coding. In these cases, features of decision-feedback adaptation are introduced into the process by which the modulation waveform estimate (or the data sequence estimate) is caused to adaptively converge on the transmitted modulation waveform (or the original
20 data sequence).

Specifically, Viterbi or other MLSE processes are applied to carefully selected elements of the modulation-waveform approximation vector $\hat{s}(n,k)$. As such, a more reliable estimate of the transmitted
25 modulation waveform and of the original data sequence is generated. Correspondingly, adaptation time is reduced. In many cases, complexity is reduced in the process of reducing the number of required adaptation iterations.

30 The two components of this method described above and respectively illustrated in FIGS. 17 and 18 may also be combined into a signal digital receiver 120'. Referring now to FIG. 19, this aspect of the

present invention includes provisions for I&Q (I and Q
sampler 192, i.e, for A/D conversion of the real and
imaginary components of the RF waveform, as well as
provisions for timing recovery 196. In this case,
5 timing recovery may be based on correlation (via
correlator 194) against an embedded reference waveform.
Timing recovery is used to drive the I&Q sampling
process as well as the timing of convergence
coefficients $\bar{\mu}_h(n,k)$ and $\bar{\mu}_s(n,k)$ used in the adaptive
10 algorithms 190, which may be included within the symbol
estimator 182, or within the channel estimator 172. A
modulator 183 and a demodulator 184 are also part of
the digital receiver 120'".

FIG. 20 illustrates another embodiment of the
15 digital receiver 120'" for simultaneously demodulating
a plurality of received digitally modulated signals
subjected to multipath propagation impairments.

The process of joint adaptation of the
channel time-response approximation $\hat{h}(n,k)$ and of the
20 transmitted modulation waveform approximation $\hat{s}(n,k)$ will
now be discussed.

Various methods may be employed to realize
practical joint adaptation. The first method of
realizing practical joint adaptation involves the
25 adaptation of the transmitted modulation-waveform
vector approximation $\hat{s}(n,k)$ simultaneously with that of
the vector channel time-response approximation $\hat{h}(n,k)$.

In a "blind" sense, $\hat{h}(n,k)$ may be initialized with a

single unit amplitude sample surrounded by all zero
amplitude samples. In a "trained" sense, the vector
channel time-response approximation $\hat{h}(n,k)$ may be
approximated through initial training based on a
5 training waveform.

The second method of realizing practical
joint adaptation comprises alternating adaptation of
large segments with respect to depth index k . For
example, the vector channel time-response approximation
10 $\hat{h}(n,k)$ is first initialized with a received training
waveform. This approximation is held constant while
the transmitted modulation-waveform vector
approximation $\hat{s}(n,k)$ is adaptively estimated over an
appropriately sized segment of samples with respect to
15 depth index k .

The size of this segment may be chosen
appropriately with respect to minimum stationary
intervals applicable to anticipated multipath. This
transmitted modulation-waveform vector approximation
20 $\hat{s}(n,k)$ is initialized in its adaptation process with the
known training waveform. At the conclusion of the
adaptive process used to converge on the transmitted
modulation-waveform vector approximation $\hat{s}(n,k)$, the
vector channel time-response approximation $\hat{h}(n,k)$
25 adaptation is resumed. The process continues back-and-
forth between adaptive convergence of $\hat{s}(n,k)$ over some

interval in domain k and subsequent vector channel
time-response approximation $\hat{h}(n,k)$ adaptation.

A third method of realizing joint adaptation
involves transformation of the modulation-waveform

5 vector-approximation recursion-equations for $\hat{s}(n,k)$ into
a single equation in one unknown variable. In other
words, linear combination or estimation is being
performed. Such an equation is formulated from the
vector channel time-response approximation $\hat{h}(n,k)$. This
10 equation is applied to known samples of $\hat{s}(n,k)$ to solve
successively for unknown samples, one at a time. The
approximation $\hat{h}(n,k)$ is updated either every time
sample n or in appropriately sized segments.

A fourth method involves the use of an
15 adaptation convergence coefficient $\mu_s(n,k)$ scaled in
magnitude over depth index k for adaptive convergence
of the modulation waveform approximation $\hat{s}(n,k)$.

All of these methods are subject to the
caveats described above. These include operation in
20 the data-sequence domain " $\hat{x}(n,k)$ " as opposed to
operation in the modulation-waveform domain " $\hat{s}(n,k)$."
These caveats also include the introduction of
"decision" activity in the approximation process in the

interest of BER performance and in the interest of reduced system complexity.

There is a significant advantage associated when operating in the data-sequence domain $\hat{\bar{x}}(n,k)$ as
5 opposed to operating in the modulation-waveform domain $\hat{\bar{s}}(n,k)$. This advantage is one of reduced complexity. This advantage is owed to the fact that, when operating in the data-sequence domain $\hat{\bar{x}}(n,k)$, the recursion equations used for adaptation need only be exercised at
10 the sample points at which data-sequence samples are present.

In summary, the use of joint modulation waveform (or data sequence) adaptation approximation and channel time-response adaptation approximation has
15 several clear advantages over conventional equalization techniques. The method of adaptive convergence on channel time-response is advantageous over adaptive convergence on inverse-channel equalization response in that adaptation is limited in time to the duration of
20 the channel time-response; a shorter convergence time is required as a consequence; required accuracy is limited to that of a fewer number of channel time-response taps as opposed to a greater number of equalizer taps otherwise necessary to accomplish
25 substantial channel-inverse filtering; and channel estimation is always mathematically realizable as opposed to inverse-channel response estimation, which is sometimes not mathematically realizable in a practical FIR filter.

30 Similarly, the use of adaptive algorithms, such as LMS, to estimate transmitted modulation

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waveforms or original data sequences is superior to MLSE methods in the following respects: there is no requirement to maintain surviving trellis paths or to calculate associated metrics; complexity does not necessarily increase with multipath delay intervals; and complexity is reduced to manageable levels in extreme cases.

Additionally, the advantages of conventional methods are applicable to the method of joint adaptive approximation of modulation waveforms or data sequences and channel time-responses. These advantages include: the ability to exploit training (data) sequences or (modulation) waveforms for improved performance as is the case for "trained" equalization; ability to initialize from a "blind" start as is the case for "blind" equalization; the ability to improve performance through "decision" processes as is the case for "decision-feedback" equalization; and the ability to improve performance through decisions based on convolutional encoding as is the case for MLSE demodulation.

FIG. 20 illustrates the extension of joint channel and modulation waveform estimation to cases where at least two modulation waveforms applied to two distinct propagation channels are received jointly. In this case, the disclosed methods apply to the reception of each modulation waveform independently through each propagation channel. The recursion equations described above are applicable subject to appropriate sub-scripting with respect to index of modulation origin (1 through N).

The received modulation waveforms are jointly recoverable under the following conditions: independent training waveforms are employed at each modulator, $s_i(n)$

through $s_n(n)$, which have sufficiently favorable autocorrelation and cross-correlation properties (near-impulse autocorrelation and very low cross-correlation); and sufficient SNR is available.

- 5 Many modifications and other embodiments of the invention will come to the mind of one skilled in the art having the benefit of the teachings presented in the foregoing descriptions and the associated drawings. Therefore, it is to be understood that the
- 10 invention is not to be limited to the specific embodiments disclosed, and that modifications and embodiments are intended to be included within the scope of the appended claims.

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